

UNITED STATES PATENT AND TRADEMARK OFFICE

BEFORE THE PATENT TRIAL AND APPEAL BOARD

SAMSUNG ELECTRONICS CO., LTD.;
SAMSUNG ELECTRONICS AMERICA, INC.,
Petitioner

v.

EMPIRE TECHNOLOGY DEVELOPMENT LLC,
Patent Owner.

Case No. IPR2024-00896
U.S. Patent No. 8,798,120

**DECLARATION OF GORDON MACPHERSON IN SUPPORT OF
PETITION FOR INTER PARTES REVIEW
OF U.S. PATENT NO. 8,798,120**

DECLARATION OF GORDON MACPHERSON

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Exhibit A	Shuguang Cui, A. J. Goldsmith and A. Bahai, "Modulation optimization under energy constraints," IEEE International Conference on Communications, 2003. ICC '03., Anchorage, AK, USA, 2003, pp. 2805-2811 vol.4, doi: 10.1109/ICC.2003.1204511.
Exhibit B	E. Tirola and J. Ylitalo, "Performance of a UMTS uplink MIMO scheme," 2003 IEEE 58th Vehicular Technology Conference. VTC 2003-Fall (IEEE Cat. No.03CH37484), Orlando, FL, USA, 2003, pp. 657-661 Vol.1, doi: 10.1109/VETECF.2003.1285099.
Exhibit C	L. Tang and M. A. Abu-Rgheff, "Joint Pilot-Aided and Blind Decision-Directed Channel Estimation for MIMO-OFDM System," 2007 IEEE 18th International Symposium on Personal, Indoor and Mobile Radio Communications, Athens, Greece, 2007, pp. 1-5, doi: 10.1109/PIMRC.2007.4394366.

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- III. Energy-Constrained MQAM
- IV. Energy-Constrained MFSK
- V. Comments on the Optimization Algorithms

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We consider radio applications where the nodes operate on batteries so that energy consumption must be minimized while satisfying given throughput and delay requirements. In this context, we analyze the best modulation strategy to minimize the total energy consumption required to send a given number of bits. The total energy consumption includes both the transmission energy and the circuit energy consumption. We show that for both MQAM and MFSK the transmission energy decreases with the $BT_{\text{sub on}}/$ product while the circuit energy consumption increases with $T_{\text{sub on}}/$, where B is the modulation bandwidth and $T_{\text{sub on}}/$ is the transmission time. Thus, in short-range applications where the circuit energy consumption is nonnegligible compared with the transmission energy, the total energy consumption is minimized by using the maximum system bandwidth along with an optimized transmission time $T_{\text{sub on}}/$. We derive this optimal $T_{\text{sub on}}/$ for MQAM and MFSK modulation in both AWGN channels and Rayleigh fading channels. Our optimization considers both delay and peak-power constraints. Numerical examples are given, where we exhibit up to 68% energy savings over modulation strategies that minimize the transmission energy alone.

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 Contents

SECTION I.

Introduction



Recent hardware advances allow more signal processing functionality to be integrated into a single chip. It is believed that soon it will be possible to integrate an RF transceiver, A/D and D/A converters, baseband processors, and other application interfaces into one device that is as small as a piece of grain and can be used as a fully-functional wireless node. Such wireless nodes typically operate with small batteries for which replacement, when possible, is very difficult and expensive. Thus, in many scenarios, the wireless nodes must operate without battery replacement for many years. Consequently, minimizing the energy consumption is a very important design consideration. In [1], the authors show that the hardware, the link layer, the MAC layer, and all other higher layers should be jointly designed to minimize the total energy consumption for energy-constrained wireless networks. The μ AMPs project [2] at MIT, the WINS project [3] at UCLA and the PicoRadio project [4] at Berkeley are investigating energy-constrained radios and their impact on overall network design.

Achieving an optimal joint design across all layers of the network protocol stack is quite challenging. We therefore consider pair-wise optimization of the hardware and link layer designs. We investigate the energy consumption associated with both the transmitting path and the receiving path: namely the total energy required to convey a given number of bits to the receiver for reliable detection. We assume that the traffic between two nodes is approximately symmetric. Thus, minimizing the total energy consumption along both the transmitting path and the receiving path is equivalent to minimizing the total energy consumption inside a given node.

The issue of energy saving is significant since in a wireless node, the battery energy is finite and hence a node can only transmit a finite number of bits. The maximum number of bits that can be sent is defined by the total battery energy divided by the required energy per bit. Most of the pioneering research in the area of energy-constrained communication has focused on transmission schemes to minimize the transmission energy per bit. In [5] the authors discuss some optimal strategies that minimize the energy per bit required for reliable transmission in the wide-band regime. In [6] the authors propose an optimal scheduling algorithm to minimize transmission energy by maximizing the transmission time for buffered packets. The emphasis on minimizing transmission energy is reasonable in the traditional wireless link where the transmission distance is large (≥ 10 m), so that the transmission energy is dominant in the total energy consumption. However, in many recently-proposed wireless ad hoc networks the nodes are densely distributed, and the average distance between nodes is usually below 10 m. In this scenario, the circuit energy consumption along the signal path becomes comparable to or even dominates the transmission energy in the total energy consumption. Thus, in order to find the optimal transmission scheme, the overall energy consumption including both transmission and circuit energy consumption needs to be considered. In [7], some insightful observations are drawn for choosing energy-efficient modulation schemes and multi-access protocols when both transmission energy and circuit energy consumption are considered. It is shown that M-ary modulation may enable energy saving over binary modulation for some short-range applications by decreasing the transmission time. In [8], MQAM modulation is analyzed in detail, and optimal strategies to minimize the total energy consumption are proposed for AWGN channels. We extended these ideas to a detailed tradeoff analysis of the transmission energy, the circuit energy consumption, the modulation bandwidth, the transmission time, and the constellation size for MQAM and MFSK in both AWGN and Rayleigh fading channels. This analysis also takes peak-power and delay constraints into account.

Modulation Optimization under Energy Constraints

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Abstract—We consider radio applications where the nodes operate on batteries so that energy consumption must be minimized while satisfying given throughput and delay requirements. In this context, we analyze the best modulation strategy to minimize the total energy consumption required to send a given number of bits. The total energy consumption includes both the transmission energy and the circuit energy consumption. We show that for both MQAM and MFSK the transmission energy decreases with the BT_{on} product while the circuit energy consumption increases with T_{on} , where B is the modulation bandwidth and T_{on} is the transmission time. Thus, in short-range applications where the circuit energy consumption is nonnegligible compared with the transmission energy, the total energy consumption is minimized by using the maximum system bandwidth along with an optimized transmission time T_{on} . We derive this optimal T_{on} for MQAM and MFSK modulation in both AWGN channels and Rayleigh fading channels. Our optimization considers both delay and peak-power constraints. Numerical examples are given, where we exhibit up to 68% energy savings over modulation strategies that minimize the transmission energy alone.

I. INTRODUCTION

Recent hardware advances allow more signal processing functionality to be integrated into a single chip. It is believed that soon it will be possible to integrate an RF transceiver, A/D and D/A converters, baseband processors, and other application interfaces into one device that is as small as a piece of grain and can be used as a fully-functional wireless node. Such wireless nodes typically operate with small batteries for which replacement, when possible, is very difficult and expensive. Thus, in many scenarios, the wireless nodes must operate without battery replacement for many years. Consequently, minimizing the energy consumption is a very important design consideration. In [1], the authors show that the hardware, the link layer, the MAC layer, and all other higher layers should be jointly designed to minimize the total energy consumption for energy-constrained wireless networks. The μ AMPs project [2] at MIT, the WINS project [3] at UCLA and the PicoRadio project [4] at Berkeley are investigating energy-constrained radios and their impact on overall network design.

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For both MQAM and MFSK we minimize the total energy consumption required to meet a given BER requirement by optimizing two system parameters: the modulation bandwidth and the transmission time. The transmission time is bounded above by the delay requirement and bounded below by the peak-power constraint. The transmission energy is analyzed via probability of error bound approximations and the circuit energy consumption is approximated as a linear function of the transmission time. We show that the total energy function has nice properties such that an optimal bandwidth-time pair can be found numerically. From this optimization, we also find the optimal constellation size for MQAM and MFSK.

The remainder of this paper is organized as follows. Section II describes the system model in detail. Section III and IV solve the energy-constrained modulation problem for MQAM and MFSK respectively, in both an AWGN channel and a Rayleigh fading channel. Section V makes some comments on the optimization algorithms. Section VI summarizes our conclusions.

II. SYSTEM MODEL

We consider a communication link connecting two wireless nodes. In order to minimize the total energy consumption, all signal processing blocks at the transmitter and the receiver

need to be considered in the optimization model. However, at this stage we neglect the energy consumption of baseband signal processing blocks (e.g., source coding, pulse-shaping, and digital modulation). We also assume that the system is uncoded. In order to clearly illustrate the tradeoff between the transmission energy and the circuit energy consumption, the power consumption of the A/D and D/A blocks is not included in the model. Otherwise, the introduced quantization noise will complicate the analysis. The methodology used here will be extended to include the A/D, D/A as well as other baseband signal processing blocks in our future research. The resulting signal paths on the transmitter and receiver sides are shown in Fig. 1 and Fig. 2, respectively. Although this model is based on a generic Low-IF transceiver structure, our framework can be easily modified to analyze other architectures as well.

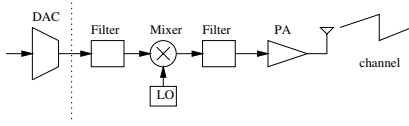


Fig. 1. Transmitter Circuit Blocks (Analog)

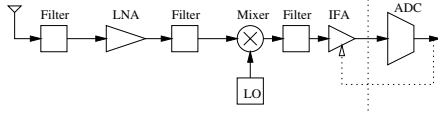


Fig. 2. Receiver Circuit Blocks (Analog)

We assume that the transceiver circuitry works on a multi-mode basis: when there is a signal to transmit all circuits work in active mode, when there is no signal to transmit they work in sleep mode, and when switching from sleep mode to active mode there is a transient mode. The multi-mode operation provides a significant saving of energy when the sleep mode is deployed. We also assume that the power consumption of the mixer, the frequency synthesizer, and the Low Noise Amplifier (LNA) can be approximated to be independent of the modulation bandwidth B , which is true when $B \ll f_c$ with f_c the carrier frequency. For the active analog filters and the Intermediate-Frequency Amplifier (IFA), their power consumption is dependent on the signal bandwidth. However, since their power consumption values are much less than that of the frequency synthesizer, the LNA or the power amplifier, they are also approximated as independent of B in our model for simplicity. Hence, as a first-order model, the circuit energy consumption can be expressed as a linear function of active time, since energy is the product of power and time.

We assume a node has L bits to transmit with a deadline T . The transceiver spends time $T_{on} \leq T$ to communicate these bits, where T_{on} is a parameter to optimize, and then returns to the sleep mode where all the circuits in the signal path are shut down for energy saving. Although the transient duration from active mode to sleep mode is short enough to be negligible, the start-up process from sleep mode to active mode may be slow due to the finite Phase Lock Loop (PLL) settling time in the frequency synthesizer. Thus, the transmission period T is given by $T = T_{tr} + T_{on} + T_{sp}$, where T_{tr} is the transient mode duration which is equal to the frequency synthesizer settling time (the start-up process of the mixer and power amplifier is

fast enough to be neglected). Correspondingly, the total energy consumption E required to send L bits also consists of three components:

$$\begin{aligned} E &= P_{on}T_{on} + P_{sp}T_{sp} + P_{tr}T_{tr} \\ &= (P_t + P_{c0})T_{on} + P_{sp}T_{sp} + P_{tr}T_{tr}, \end{aligned} \quad (1)$$

where P_{on} , P_{sp} and P_{tr} are power consumption values for the active mode, the sleep mode, and the transient mode, respectively. The active mode power P_{on} comprises the transmission signal power P_t and the circuit power consumption P_{c0} in the whole signal path. Specifically, P_{c0} consists of the power amplifier power consumption P_{amp} , the mixer power consumption P_{mix} , the frequency synthesizer power consumption P_{syn} , the LNA power consumption P_{LNA} , the active filter power consumption P_{fil} (including all filters in the signal path), and the IFA power consumption P_{IFA} , where $P_{amp} = \alpha P_t$ and $\eta = \frac{1}{1+\alpha}$ is the drain efficiency [10] of the RF power amplifier. Although strictly speaking P_t should be part of the total amplifier power consumption, here we define P_{amp} as the value excluding the transmission signal power for convenience.

Since batteries are not only energy-limited but also peak-power-limited, the total power consumption of the wireless node can never exceed the maximum battery output power. The maximum power available for transmission and the related circuits, denoted as P_{max} , is equal to the maximum battery output power minus the total power consumption in all other unaccounted blocks. Since $P_{on} = \max\{P_{on}, P_{tr}, P_{sp}\}$, the peak-power constraint is given by

$$\begin{aligned} P_{on} &= P_t + P_{amp} + P_c \\ &= (1 + \alpha)P_t + P_c \\ &\leq P_{max}, \end{aligned} \quad (2)$$

where $P_c = 2P_{mix} + 2P_{syn} + P_{LNA} + P_{fil} + P_{IFA}$ is used to denote the circuit power consumption (excluding the power amplifier power consumption) in the active mode and $P_c T_{on}$ will be used to denote the corresponding circuit energy consumption in the following sections.

The start-up time for other circuit blocks is negligible compared to that of the frequency synthesizers. Hence, the optimal strategy for the start-up process is to turn on the frequency synthesizers first, and once they settle down, turn on the rest of the circuits. As a result, there is no unnecessary energy-wasting while the transceiver waits for the frequency synthesizers to settle down. Hence, P_{tr} merely needs to include the power consumption of the frequency synthesizers.

In the sleep mode, the power consumption is dominated by the leaking current I_l of the switching transistors if the circuitry is properly designed. Since CMOS technology is exclusively used in our analysis, the leaking current $I_l \approx 0$ [11] compared with the active mode. Thus, we set $P_{sp} = 0$. Our analysis can be easily modified to incorporate $P_{sp} \neq 0$.

Given Eq. (1) and Eq. (2), and the fact $P_{sp} = 0$ and $P_{tr} \approx 2P_{syn}$, the energy consumption per information bit $E_a = E/L$ is given by

$$\begin{aligned} E_a &= ((1 + \alpha)P_t + P_c)T_{on} + P_{tr}T_{tr})/L \\ &\approx ((1 + \alpha)E_t + P_c T_{on} + 2P_{syn}T_{tr})/L, \end{aligned} \quad (3)$$

where $E_t = P_t T_{on}$, and P_c , P_{syn} and T_{tr} can be treated as constants defined by the particular transceiver structure in use. As shown in the following sections, the transmission energy

E_t is a monotonically increasing function of the bandwidth efficiency defined as $B_e = L/(BT_{on})$ (in bits/s/Hz). In other words, E_t is a monotonically decreasing function of the BT_{on} product for any fixed packet size L .

One thing we need to point out is that since the model shown in Fig. 1 and Fig. 2 is a generic model, it may need some modifications for specific systems. For example, the mixer at the transmitter side needs to be deleted for MFSK systems since frequency modulation is usually implemented digitally inside the frequency synthesizer. As a result, the corresponding expression for the P_c term in all the energy consumption formulas needs to be revised.

Finally, the energy-constrained modulation problem can be modeled as

$$\begin{aligned} & \text{minimize} && E_a \\ & \text{subject to} && 0 \leq T_{on} \leq T - T_{tr} \\ & && 0 \leq B \leq B_{max} \\ & && 0 \leq (1 + \alpha)P_t + P_c \leq P_{max} \end{aligned} \quad (4)$$

where we can see that our task is to find the most efficient way to use the available bandwidth and transmission time under the given system constraints so that the total energy consumption is minimized. From the resulting optimal (B, T_{on}) pair, the optimal constellation size for a particular modulation scheme can be obtained. In the following sections, the optimization problem is solved for both MQAM and MFSK.

III. ENERGY-CONSTRAINED MQAM

We first take MQAM as a design example. Analysis is done over an AWGN channel and a Rayleigh fading channel, respectively.

A. AWGN Channels

For MQAM, the number of bits per symbol is defined as $b = \log_2 M$. The number of MQAM symbols needed to send L bits is denoted as $L_s = \frac{L}{b}$. If the symbol period is denoted as T_s , we can also represent L_s as $L_s = \frac{T_{on}}{T_s}$. Thus, $\frac{L}{b} = \frac{T_{on}}{T_s}$, i.e.,

$$b = \frac{LT_s}{T_{on}}. \quad (5)$$

If square pulses are used, $T_s \approx 1/B$. Hence,

$$b \approx \frac{L}{BT_{on}}. \quad (6)$$

Since the bandwidth efficiency is defined as $B_e = \frac{L}{BT_{on}}$, we can see that $b \approx B_e$ for MQAM.

The probability of error bound for MQAM is given by [12]

$$P_e \leq 4(1 - \frac{1}{2^b})Q\left(\sqrt{\frac{3}{2^b - 1}}\gamma\right) \leq 4(1 - \frac{1}{2^b})e^{-\frac{3}{2^b - 1}\frac{\gamma}{2}}, \quad (7)$$

where $Q(x) = \int_x^\infty \frac{1}{\sqrt{2\pi}} e^{-\frac{u^2}{2}} du$. The Signal to Noise Ratio (SNR) γ is defined as $\gamma = \frac{P_r}{2B\sigma^2 N_f}$, where P_r is the received signal power, σ^2 is the power spectral density of the AWGN and N_f is the receiver noise figure defined as $N_f = \frac{N_{total}}{2B\sigma^2}$,

where N_{total} is the total noise power which combines the channel noise and the noise introduced by all receiver circuits.

Hence, by approximating the bound as equality we obtain

$$P_r \approx \frac{4}{3} N_f B \sigma^2 (2^b - 1) \ln \frac{4(1 - \frac{1}{2^b})}{P_e}. \quad (8)$$

Assuming free space propagation at distance d (meters), the transmission power is equal to

$$P_t = P_r G_d, \quad (9)$$

with $G_d \triangleq \frac{(4\pi d)^2}{G\lambda^2}$ the power gain factor, λ the carrier wavelength, and G a constant defined by the antenna gain and other system parameters. According to Eq. (6), Eq. (8) and Eq. (9), we obtain the transmission energy as

$$\begin{aligned} E_t &= P_t T_{on} \\ &\approx \frac{4}{3} N_f \sigma^2 (2^{\frac{L}{BT_{on}}} - 1) \ln \frac{4(1 - 2^{-\frac{L}{BT_{on}}})}{P_e} G_d B T_{on}. \end{aligned} \quad (10)$$

It is easily shown that E_t is a monotonically decreasing function over the product BT_{on} when MQAM is well defined, i.e., when $b = \frac{L}{BT_{on}} \geq 2$. Thus, when the packet size L is fixed, there are two ways to decrease the transmission energy: increase the bandwidth B or increase the time duration T_{on} . Since the value of E_t is only dependent on the product of B and T_{on} , the optimal bandwidth-time pair to minimize the transmission energy will always be the maximum B together with the maximum T_{on} .

However, when we include the circuit energy consumption in the model, the situation may change. According to Eq. (3), the expression for the total energy consumption per information bit in term of T_{on} is given by

$$\begin{aligned} E_a &= ((1 + \alpha) \frac{4}{3} N_f \sigma^2 (2^{\frac{L}{BT_{on}}} - 1) \ln \frac{4(1 - 2^{-\frac{L}{BT_{on}}})}{P_e} \\ &\quad \times G_d B T_{on} + P_c T_{on} + 2P_{syn} T_{tr}) / L, \end{aligned} \quad (11)$$

where we see that in order to minimize the total energy consumption we must use the maximum bandwidth since the circuit energy consumption is independent of B and the transmission energy decreases with B . However, the value for T_{on} needs to be optimized under the delay and peak-power constraints. The peak-power constraint in Eq. (4) can be rewritten as

$$\begin{aligned} (1 + \alpha) \frac{4}{3} N_f \sigma^2 (2^{\frac{L}{BT_{on}}} - 1) \ln \frac{4(1 - 2^{-\frac{L}{BT_{on}}})}{P_e} G_d B \\ \leq P_{max} - P_c. \end{aligned} \quad (12)$$

For $B = B_{max}$, this inequality is equivalent to

$$T_{on} \geq T_{min}, \quad (13)$$

where T_{min} is the solution for T_{on} for which the equality in Eq. (12) holds. Thus, for MQAM the optimization model in Eq. (4) can be rewritten as

$$\begin{aligned} & \text{minimize} && E_a \\ & \text{subject to} && T_{min} \leq T_{on} \leq T - T_{tr} \\ & && B = B_{max} \end{aligned} \quad (14)$$

$f_c = 2.5$ GHz	$\alpha = 1.9$
$G = 1$	$\sigma^2 = \frac{N_0}{2} = 10^{-16}$ W/Hz
$B_{max} = 1$ MHz	$L = 200$ kb
$P_{mix} = 30.3$ mW	$P_{syn} = 50$ mW
$P_{LNA} = 20$ mW	$P_{IFA} = 3$ mW
$P_{max} = 250$ mW	$P_{fil} = 5$ mW
$T_{tr} = 5$ μ s	$T_s = \frac{1}{B}$
$T = 100$ ms	$P_e = 10^{-3}$
$N_f = 10$ dB	

TABLE I
MQAM PARAMETERS

For a specific numerical example, the circuit-related parameters need to be defined first. Here, the associated circuitry is composed of several blocks quoted from [10], [13], [14], and [15]. The corresponding parameters are summarized in Table I, where $\alpha = 1.9$ corresponds to the drain efficiency $\eta = 0.35$, which is a practical value for class-A RF power amplifiers [10] (Due to the linearity requirement for amplifying MQAM signals, class-A power amplifiers are usually used.). The values for B_{max} , L , and T are set up in such a way that $b_{min} = \frac{L}{B_{max}T} = 2$. Thus, the constellation size for MQAM is well defined inside the feasible region.

The plot of E_a over T_{on} for a maximum bandwidth $B = 1$ MHz is shown in Fig. 3. The vertical axis is the energy consumption per information bit (in terms of dB relative to a millijoule: $\log_{10} \frac{E_a}{0.001}$ dBmJ). The horizontal axis is the normalized transmission time. We see that the total energy consumption is no longer a monotonically-decreasing function over T_{on} when the transmission distance d is small. For example, when $d = 1$ m, E_a at the optimal $T_{on}^* \approx 0.28T$ is about 5 dB lower than the non-optimized case where $T_{on} = T - T_{tr}$. Thus, T_{on} optimization results in a 68% energy saving. To guarantee that the peak-power constraint is satisfied, we need to verify that T_{on}^* is larger than T_{min} . From Eq. (12) we can numerically verify that the peak-power constraints are satisfied when $d \leq 10$ m. If $T_{on}^* < T_{min}$ does occur, $T_{on} = T_{min}$ is chosen as the final solution. For example, when P_{max} is decreased from 250 mW to 189 mW, $T_{min} = 0.41T$ when $d = 1$ m. Thus, $T_{on} = 0.41T$ is used instead of $T_{on} = 0.28T$ for $d = 1$ m.

The transmission energy is dependent on the transmission distance d while obviously the circuit energy consumption is independent of d . Thus, we can save energy by optimizing T_{on} only when the circuit energy consumption is nontrivial relative to the transmission energy. Since the transmission energy increases with d , there exists a threshold for the value of d above which there is no energy savings possible. For the above example, $d = 10$ m is the threshold, where the derivative of E_a relative to T_{on} is approximately zero at the point $T_{on}/T \approx 1$. In general, to find the threshold we just need to find the value of d that makes the derivative of E_a relative to T_{on} at the maximum transmission time equal to zero.

According to the relationship defined in Eq. (6), we can find the optimal constellation size from the optimal values of B and T_{on} . We redraw E_a over b for the $d = 1$ m case in Fig. 4. We see from the figure that $b_{opt} \approx 7$ if the total energy consumption is considered versus $b_{opt} = 2$, its minimum value, when only transmission energy is considered.

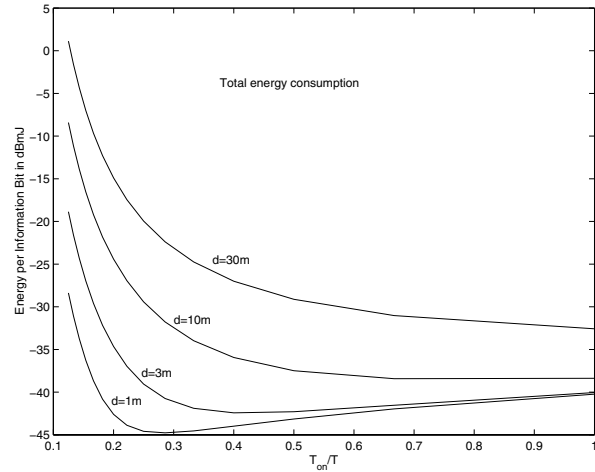


Fig. 3. Total Energy Consumption, MQAM, (AWGN)

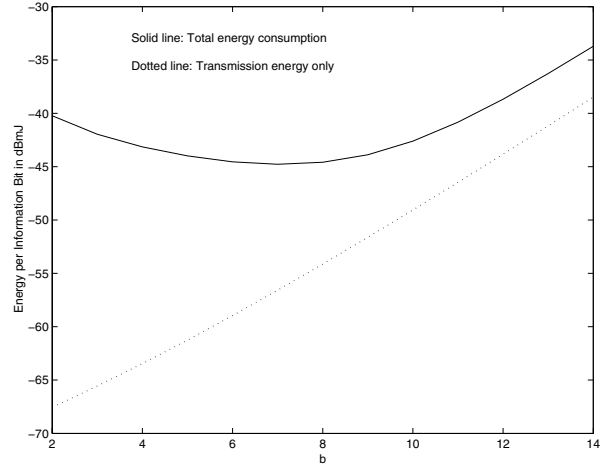


Fig. 4. Total energy consumption versus constellation size, MQAM (AWGN)

B. Rayleigh Fading Channels

In a flat Rayleigh fading channel, the average probability of error is given by (see [16])

$$P_e \approx \frac{1}{2} \left(1 - \sqrt{\frac{g\bar{\gamma}}{1 + g\bar{\gamma}}} \right), \quad (15)$$

where $g = \frac{3}{2(M-1)}$ for MQAM, and the average SNR $\bar{\gamma}$ is defined as $\bar{\gamma} = \int \gamma p(\gamma) d\gamma$, where $p(\gamma) = \frac{1}{\bar{\gamma}} e^{-\gamma/\bar{\gamma}}$ is the SNR distribution in a Rayleigh fading channel.

From Eq. (15), we derive

$$\bar{\gamma} \approx \frac{(1 - 2P_e)^2}{g(1 - (1 - 2P_e)^2)} \approx \frac{1}{4gP_e} \approx \frac{1}{4 \frac{3}{2(M-1)} P_e} = \frac{2^b - 1}{6P_e}, \quad (16)$$

where we used the fact that $P_e \ll 1$. Considering $\bar{\gamma} = \frac{P_r}{2B\sigma^2 N_f}$, the required average received power for the target average

probability of error P_e is given by

$$P_r \approx N_f B \sigma^2 \frac{2^b - 1}{3P_e} = N_f B \sigma^2 \frac{2^{L/(BT_{on})} - 1}{3P_e}. \quad (17)$$

We still assume that the average signal power is attenuated according to the square-law. Thus, the required transmission power for the target average probability of error P_e is given by

$$P_t = P_r G_d = N_f B \sigma^2 \frac{2^{L/(BT_{on})} - 1}{3P_e} G_d. \quad (18)$$

The energy consumption per information bit at the transmitter is expressed as

$$E_a = \left((1 + \alpha) N_f \sigma^2 \frac{2^{L/(BT_{on})} - 1}{3P_e} G_d B T_{on} + P_c T_{on} + 2P_{syn} T_{tr} \right) / L. \quad (19)$$

The optimization model can also be simplified into the form of Eq. (14), where the value of T_{min} is calculated in the same way as in AWGN channels. For numerical examples, we use the same setup as described in Table I. The energy consumption over T_{on} is plotted in Fig.5. Since in a fading channel the transmission energy requirement is much higher than that in an AWGN channel with the same BER requirement, the transmission energy consumption becomes dominant more quickly when we decrease T_{on} . Hence, the energy we can save by choosing the optimal T_{on} is less than that in an AWGN channel at the same distance. Non-trivial energy savings only occurs when the transmission distance is very small (e.g., $d = 0.1m$).

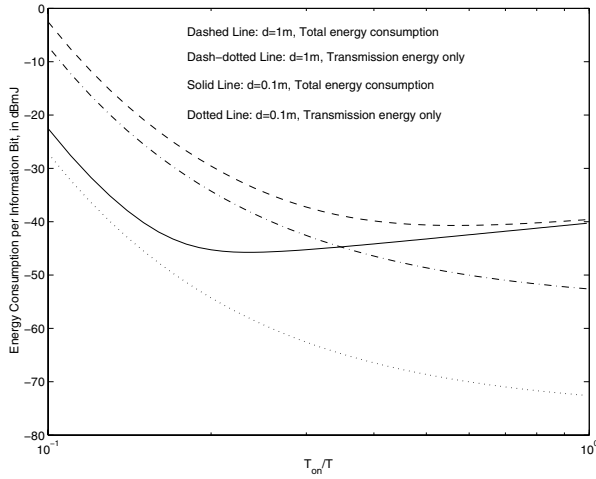


Fig. 5. Total energy consumption versus T_{on} , MQAM (Rayleigh fading)

IV. ENERGY-CONSTRAINED MFSK

For MFSK, the number of orthogonal carriers is $M = 2^b$. We assume the carrier separation is equal to $\frac{1}{2T_s}$, where T_s is the symbol period. Thus, the data rate $R = \frac{b}{T_s}$ and the total

bandwidth can be approximated as $B \approx \frac{2^b}{2T_s}$ (see [12]). As a result, the bandwidth efficiency for MFSK is given by

$$B_e \triangleq 2b/2^b \text{ (b/s/Hz)}. \quad (20)$$

Since in general the bandwidth efficiency can also be represented as $B_e = \frac{L}{BT_{on}}$ regardless of modulation schemes, the relationship between the constellation size and the bandwidth-time product for MFSK is given by

$$2b/2^b = L/(BT_{on}), \quad (21)$$

which is different from the MQAM case where $b = L/(BT_{on})$. However, there still exists a one-to-one relationship between b and the BT_{on} product for any fixed value of L , except for $b = 1$ and $b = 2$, which correspond to the same BT_{on} product. We will discuss this interesting phenomena at the end of this section. For simplicity, all the following transmission energy functions for MFSK will be represented in term of b .

A. AWGN Channel

Since most practical MFSK receivers use non-coherent detectors, the probability of error bound for non-coherent MFSK detection is used in our derivation:

$$P_e \leq \frac{2^b - 1}{2} e^{-\frac{\gamma}{2}}. \quad (22)$$

Approximating the bound as an equality, we obtain $\gamma = \frac{bE_b}{N_0 N_f} \approx 2 \ln \frac{2^b - 1}{2P_e}$, where E_b is the energy per information bit at the receiver and $N_0 = 2\sigma^2$. Hence, $\frac{E_b}{2\sigma^2 N_f} = \frac{2}{b} \ln \frac{2^b - 1}{2P_e}$. Following the same way as in the MQAM case, the transmission power and the transmission energy are given by

$$P_t = 4N_f \sigma^2 \ln \frac{2^b - 1}{2P_e} G_d \frac{2B}{2^b} \quad (23)$$

and

$$E_t = P_t T_{on} = 4N_f \sigma^2 \ln \frac{2^b - 1}{2P_e} G_d \frac{L}{b} \quad (24)$$

respectively, where we used the fact $T_{on} = \frac{2^b}{2b} \frac{L}{B}$ derived from Eq. (21).

One necessary modification in the hardware configuration compared to the MQAM system is that the mixer at the transmitter side should be deleted as we discussed in Section II. Correspondingly we redefine $P_c = 2P_{syn} + P_{mix} + P_{LNA} + P_{fil} + P_{IFA}$. Thus, the total energy consumption per information bit is given by

$$E_a = \left((1 + \alpha) 4N_f \sigma^2 \ln \frac{2^b - 1}{2P_e} G_d \frac{L}{b} + P_c T_{on} + 2P_{syn} T_{tr} \right) / L. \quad (25)$$

From Eq. (21) we see that the product of B and T_{on} defines the value of b when the packet size L is fixed. Hence, the value of E_t is dependent on the BT_{on} product. It can be proved that E_t is a monotonically decreasing function over b (when $b \geq 1$) unless P_e is unreasonably large (on the order of 0.1). Due to

the relationship between b and BT_{on} as described in Eq. (21), E_t is also a monotonically decreasing function over the BT_{on} product which is similar to the MQAM case. Hence, increasing B always decreases the total energy consumption, but the optimal T_{on} may not be the maximum allowable transmission time. The optimization model is easily described in term of b . When $B = B_{max}$, it turns out that Eq. (4) can be rewritten as

$$\begin{aligned} & \text{minimize} && E_a \\ & \text{subject to} && b_{min} \leq b \leq b_{max} \\ & && B = B_{max} \end{aligned} \quad (26)$$

where b_{min} is calculated based the peak-power constraint, specifically, $(1 + \alpha)P_t(b_{min}) = P_{max} - P_c$, and b_{max} is defined by the delay requirement in such a way that $\frac{2b_{max}}{2^{b_{max}}} = \frac{L}{B_{max}T}$.

To give a numerical example, we first assume the power consumption of corresponding circuit blocks is roughly the same as in the MQAM case. Since there is no longer a strict linearity requirement on the RF power amplifier, the value of α in Table I is changed to 0.33, i.e., the drain efficiency $\eta = .75$, which corresponds to a class-B or a higher-class (C,D or E) power amplifier. The maximum bandwidth B_{max} and the packet size L are kept as 1 MHz and 200 kb, respectively. The maximum delay T is changed to 1.07 s, such that $b_{max} = 6$. Compared with the MQAM case, the increment of T is due to the fact that MFSK is less bandwidth-efficient than MQAM.

Hence, by fixing $B = B_{max} = 1$ MHz, we draw E_t and E_a directly over b as shown in Fig. 6. Not surprisingly, the

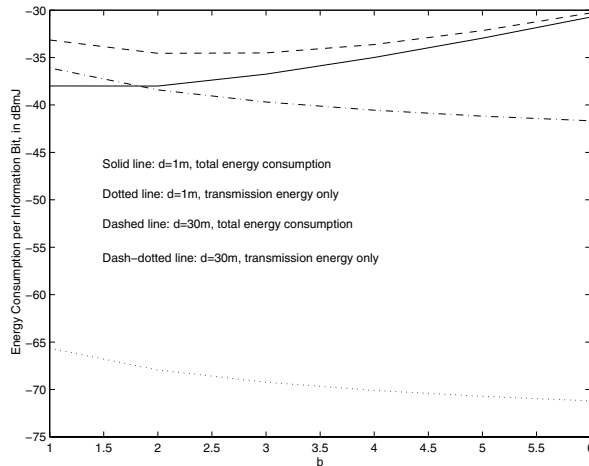


Fig. 6. Total energy consumption versus b , MFSK (AWGN)

transmission energy E_t goes down when b increases, since it is well known that the larger M is, the more energy efficient MFSK is, in an AWGN channel. In other words, $M = \infty$ is optimal in the sense of minimizing the energy consumption per information bit ([12]). When the circuit energy consumption is considered as shown in Fig. 6, 4-FSK turns out to be the best choice for both $d = 1$ m and $d = 30$ m if we do not consider the peak-power constraint. To take the peak-power constraint into account, we can numerically obtain $b_{min} = 1$ and $b_{min} = 4$ for $d = 1$ m and $d = 30$ m respectively. Thus, the final optimal values for b are $b_{opt} = 2$ for $d = 1$ m and $b_{opt} = 4$ for $d = 30$ m, respectively.

B. Rayleigh Fading Channel

Using the tight bounds derived in Chapter 7 of [12] for the P_e of MFSK in AWGN:

$$P_e \leq \begin{cases} 2e^{-b(\frac{\gamma}{4} - 2 \ln 2)/2} & (\ln M \leq \frac{\gamma}{4}) \\ 2e^{-b(\sqrt{\frac{\gamma}{b}} - \sqrt{\ln 2})^2} & (\frac{\gamma}{4} \leq \ln M \leq \gamma) \end{cases} \quad (27)$$

and taking the average over the distribution of the received power, we can numerically derive a new bound for the average probability of error in a Rayleigh fading channel. From this new bound we can numerically calculate the relationship between E_b and b for any given average P_e . Then the total energy consumption is calculated as

$$E_a = (1 + \alpha)N_f E_b G_d + (P_c T_{on} + 2P_{syn} T_{tr})/L, \quad (28)$$

where $T_{on} = \frac{2^b}{2b} \frac{L}{B}$. For a numerical example, we use the same setup as in the AWGN channel case. The transmission energy and the total energy consumption are drawn in Fig. 7. We see that $b_{opt} = 4$ and $b_{opt} = 2$ for $d = 5$ m and $d = 1$ m, respectively, if the total transmission energy is considered. We can numerically check that the peak-power constraint is satisfied when $P_{max} = 250$ mW.

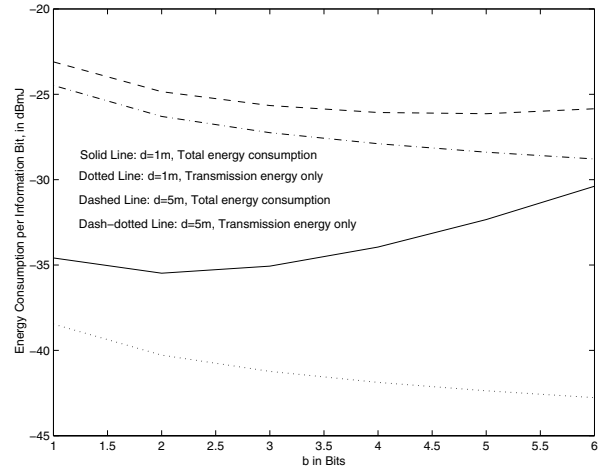


Fig. 7. Total energy consumption versus b , MFSK (Rayleigh fading)

One interesting observation is that for both AWGN and Rayleigh fading channels, 4-FSK is always more energy-efficient than BFSK. This is because the transmission energy is monotonically decreasing over b , while the circuit energy consumption is the same for the two cases since $T_{on} = \frac{2^b}{2b} \frac{L}{B}$ is the same for both $b = 1$ and $b = 2$. This result is based on the assumption that both BFSK and 4-FSK use the same transceiver architecture.

Another interesting observation is that for MFSK, there are two ways to increase b in order to decrease the transmission energy. According to the relationship $\frac{2b}{2^b} = \frac{L}{BT_{on}}$, to increase b , we can either increase B or increase T_{on} for a fixed L . The first approach is achieved by exponentially expanding the bandwidth ($B = \frac{2^b}{2T_s}$) while keeping the carrier separation

$\frac{1}{2T_e}$ constant, and the second approach is achieved by packing a larger number of orthogonal carriers inside the fixed bandwidth as a result of exponentially decreasing the carrier separation (then the T_{on} is increased accordingly since $T_{on} = \frac{2^b L}{2b}$). The two strategies work equivalently well since the product $BT_{on} = \frac{2^b}{2b}L$ (according to Eq. (21)) is the same when the value of b is the same for these two approaches. Thus, the wide-band optimality (see [12]) of MFSK in the sense of saving transmission energy can be achieved by utilizing either infinite bandwidth or infinite transmission time.

V. COMMENTS ON THE OPTIMIZATION ALGORITHMS

For MQAM, the transmission energy E_t can be shown to be monotonically decreasing over the BT_{on} product inside the feasible region ($b \geq 2$) in both AWGN and Rayleigh fading channels. Since $b = \frac{L}{BT_{on}}$ can only take on discrete values for an uncoded system, it is not useful to talk about convexity. Simple searching algorithms can be used to find the optimal solution due to the monotonicity of E_t and the fact that the feasible region is finite in practical systems.

For MFSK, monotonicity of E_t over b can be shown numerically unless the probability of error P_e takes extremely large values (on the order of 0.1). This dependence on P_e is due to the looseness of the probability of error bounds we use. As in the MQAM case, simple searching algorithms can be used to find the optimal solution.

VI. CONCLUSIONS

We have shown that for transmitting a given number of bits, the traditional belief that a longer transmission duration lowers energy consumption may be misleading if the circuit energy consumption is included, especially for short-range applications. For both MQAM and MFSK, we show that the transmission energy is completely dependent on the product of B and T_{on} . To minimize the transmission energy, both maximum bandwidth and maximum transmission time are desired. To minimize the total energy consumption, the maximum bandwidth is still desired and the transmission time needs to be optimized under the delay and peak-power constraints. The optimization is done for both AWGN and Rayleigh fading channels with up to 80% energy savings. We show that at the same transmission distance, the possible energy saving by optimizing the transmission time in a Rayleigh fading channel is less than that in an AWGN channel, since the transmission energy is more dominant in the former channel.

The energy-constrained modulation results can be applied to the optimal transmission adaptation problem to minimize the energy consumption. We assume that the transmitter works in a periodic manner, where the period T is defined by the delay requirement. At the beginning of each period T , the energy-constrained modulation problem is solved according to the number of bits in the buffer to be transmitted. In order to inform the receiver of the constellation size update, a short signaling header is transmitted to the receiver coded with a pre-defined system setup. Thus, an energy-minimizing adaptation is achieved.

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EXHIBIT B



All

▼ Joint Pilot-Aided and Blind Decision-Directed Channel Estimation for MIMO



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In this study, uplink (UL) performance was studied in the 3rd generation UMTS (universal mobile telecommunications system) system assuming different SIMO/MIMO (single/multiple input multiple output) approaches. Link simulations and theoretical analysis were used to investigate the uplink capacity and coverage of the considered antenna configurations. Link simulations assumed that the user equipment (UE) has a maximum of two transmit antennas while the base station can employ up to four antennas. The considered MIMO scheme can be classified as diversity MIMO, which means that the UE transmits the same data symbols via two antennas. Results of the performance evaluation demonstrate that receive diversity is a dominant factor in improving uplink capacity and coverage. For example, in terms of isolated cell capacity, a MIMO technique with 2 transmit (Tx) and 2 receive (Rx) antennas performed approximately 50% worse than a traditional approach with 1 Tx and 4 Rx antennas. Results show also that MIMO transmission will provide some benefit (up to 2 dB) for UL coverage and UE battery consumption. However, doubling the number of receiving antennas is still much more efficient than introducing the 2nd generation transmit antenna at the UE. Based on this study, the MIMO concept does not seem feasible for improving UMTS uplink performance.

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 Contents

SECTION I.

Introduction



Multimedia messaging is expected to dominate wireless communications in near future, which demands enhanced data throughput for both uplink and downlink directions. It is well acknowledged that improved data rates with good spectral efficiency can be achieved by increasing the number of transmit and receive antennas [1]. This MIMO approach is still in a starting phase in different standardization bodies for radio communications.

As videophones become more and more popular, it is extremely important to reach high spectral efficiency also for the uplink direction. The goal is to achieve reasonably high data rates with limited transmission power of the user equipment. Enhanced uplink performance is crucial for reducing the UE transmission power requirements, especially for high data rate applications. In addition to reduced UE battery consumption, an improved UL performance translates into better coverage area for the high data rate services [2], [3], [4], [5]. On the other hand increasing the uplink capacity and coverage reduces the number of base station sites and the overall cost of the network. Thus the performance improvement obtained using advanced antenna techniques including MIMO can bring benefit for both the network operator and the end user of a mobile terminal.

High spectral efficiency in uplink can be obtained via multiple antennas combined with link adaptation techniques. In FDD (Frequency-Division Duplex) systems this requires a reference channel in the uplink and a feedback channel in the downlink. Since the feedback rate is limited by the downlink traffic requirements only open loop methods were chosen for the current study. Typically, in code division multiple access systems uplink data transmission can be characterized by the following key factors: limited pilot power, no strict code limitation, and diversity reception at base station.

The paper is organized as follows. Section II introduces the considered transmission/reception schemes and describes the main assumptions for the performance evaluation. Section III illustrates the simulated link performance results for single user scenario. In Section IV the UL capacity results for isolated cell are presented. Section V presents and discusses the UL coverage results and finally, concluding remarks will be given in Section VI.

SECTION II.

Performance Evaluation

Fig. 1 illustrates block diagram of the considered SIMO [6] and MIMO transmission schemes. In the MIMO scheme under study the UE transmits the same data streams (both the Dedicated Physical Data Channel, DPDCH and the Dedicated Physical Control Channel, DPCCH) simultaneously via two antennas. In order to separate the two chip streams at the Node B the two transmit antennas use different channelization codes. Thus, the considered MIMO scheme doubles the utilization of the UL channelization codes when compared to the single antenna transmission or STTD [7] (Space

Performance of a UMTS Uplink MIMO Scheme

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Abstract— In this study uplink (UL) performance was studied in the 3rd generation UMTS (Universal Mobile Telecommunications System) system assuming different SIMO/MIMO (Single/Multiple Input Multiple Output) approaches. Link simulations and theoretical analysis were used to investigate the uplink capacity and coverage of the considered antenna configurations. Link simulations assumed that the user equipment (UE) has a maximum of two transmit antennas while the base station can employ up to four antennas. The considered MIMO scheme can be classified as diversity MIMO, which means that the UE transmits the same data symbols via two antennas. Results of the performance evaluation demonstrate that receive diversity is a dominant factor in improving uplink capacity and coverage. For example, in terms of isolated cell capacity, a MIMO technique with 2 transmit (Tx) and 2 receive (Rx) antennas performed approximately 50% worse than a traditional approach with 1 Tx and 4 Rx antennas. Results show also that MIMO transmission will provide some benefit (up to 2 dB) for UL coverage and UE battery consumption. However, doubling the number of receiving antennas is still much more efficient than introducing the 2nd transmit antenna at the UE. Based on this study the MIMO concept does not seem feasible for improving UMTS uplink performance.

I. INTRODUCTION

Multimedia messaging is expected to dominate wireless communications in near future, which demands enhanced data throughput for both uplink and downlink directions. It is well acknowledged that improved data rates with good spectral efficiency can be achieved by increasing the number of transmit and receive antennas [1]. This MIMO approach is still in a starting phase in different standardization bodies for radio communications.

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improvement obtained using advanced antenna techniques including MIMO can bring benefit for both the network operator and the end user of a mobile terminal.

High spectral efficiency in uplink can be obtained via multiple antennas combined with link adaptation techniques. In FDD (Frequency-Division Duplex) systems this requires a reference channel in the uplink and a feedback channel in the downlink. Since the feedback rate is limited by the downlink traffic requirements only open loop methods were chosen for the current study. Typically, in code division multiple access systems uplink data transmission can be characterized by the following key factors: limited pilot power, no strict code limitation, and diversity reception at base station.

The paper is organized as follows. Section II introduces the considered transmission/reception schemes and describes the main assumptions for the performance evaluation. Section III illustrates the simulated link performance results for single user scenario. In Section IV the UL capacity results for isolated cell are presented. Section V presents and discusses the UL coverage results and finally, concluding remarks will be given in Section VI.

II. PERFORMANCE EVALUATION

Fig. 1 illustrates block diagram of the considered SIMO [6] and MIMO transmission schemes. In the MIMO scheme under study the UE transmits the same data streams (both the Dedicated Physical Data Channel, DPDCH and the Dedicated Physical Control Channel, DPCCH) simultaneously via two antennas. In order to separate the two chip streams at the Node B the two transmit antennas use different channelization codes. Thus, the considered MIMO scheme doubles the utilization of the UL channelization codes when compared to the single antenna transmission or STTD [7] (Space Time Transmit Diversity, used in 3GPP downlink) but without the need for STTD encoder/decoder. The link performance of the current MIMO scheme is the same as with the UL STTD since both schemes represent two-way transmit diversity.

The considered base station receiver performs MRC (Maximal Ratio Combining) over signal multipath components, antennas, and Tx-diversity branches (standard Rake receiver). Doubling the number of transmit/receive antennas will double the base band complexity of the considered base station receiver.

Five different antenna configurations are considered in the link simulations: 1x1 (one transmit antenna, one receive antenna), 1x2, 2x2, 1x4 and 2x4, respectively. In all the cases both transmit and receive antennas are assumed to be uncorrelated. Furthermore, the Branch Power Ratio (BPR) of the multi element array was assumed to be zero dB [8]. Table I illustrates the temporal properties of the considered radio channel models.

The link performance of different antenna configurations is evaluated by means of Monte-Carlo simulations. The main simulation parameters and assumptions are shown in Table II. Full 3GPP (3rd Generation Partnership Project) link level modeling was used with inner and outer loop power control and realistic channel and interference estimation algorithms. Circuit-switched 64 kb/s data service specified in 3GPP conformance tests [9] was used in the link simulations. The service related parameters are summarized in Table III.

When studying the UL performance of different transmission schemes, the UL load factor, denoted as η plays an important role. The UL load factor is calculated as [2]

$$\eta = (1 + i) \cdot \sum_{j=1}^J \frac{1}{1 + \frac{(E_c / N_0)_j}{1}} \quad (1)$$

where i is the ratio of other cell to own cell interference, J is the number of users and E_c/N_0 is the chip energy per total noise power density. E_c/N_0 is calculated from the simulated Rx E_b/N_0 as

$$(E_c / N_0)_j = \text{user bit rate} \cdot (\text{Rx } E_b / N_0) \cdot \text{chip rate} \quad (2)$$

Setting $i = 0$ in (1) gives the load factor in an isolated cell, in which case the amount of other cell interference is zero. Noise rise is used as a final performance criterion for different transmission schemes. It is calculated from the UL load factor as

$$\text{Noise_rise} = 10 * \log_{10} \left(\frac{1}{1 - \eta} \right) \quad (3)$$

Equations (1) and (2) can be used to calculate the noise rise based on the simulated single user Rx E_b/N_0 values. In the performance evaluation we use another approach. As mentioned, we are doing multi user simulations, which give us Rx E_b/N_0 results in case of multiple simultaneous users. We calculate the noise rise in case of j users as

$$\text{Noise_rise}(j) = \text{Rx } \frac{E_b}{N_0}(j) - \text{Rx } \frac{E_b}{N_0}(1) + \text{Noise_rise}(1) \quad (4)$$

where Rx $E_b/N_0(j)$ (dB) is the simulated average Rx E_b/N_0 value (dB) obtained from the multi-user simulation having j simultaneous users and Noise_rise(1) (dB) corresponds the noise rise caused by a single user and it is calculated using (3) and (4).

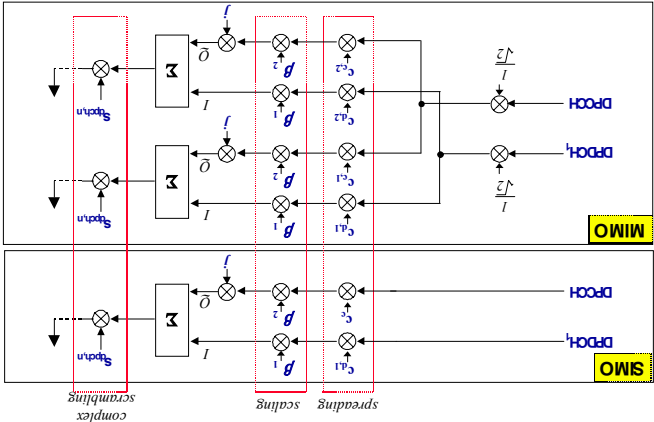


Figure 1. Block diagram of the considered transmission schemes.

TABLE I.
CHANNEL MODELS USED IN PERFORMANCE SIMULATIONS

Channel model	Relative path powers [dB]	UE velocity [km/h]
Flat fading (1-path)	0	3, 50
modified ITU Pedestrian A (2-paths)	0 - 12.7	3
modified ITU Vehicular A (6-paths)	0 -1.9 -7.3 -10.4 -10.9 -17.3	50

TABLE II.
SIMULATION PARAMETERS

Carrier frequency	1940 MHz
Chip rate	3.840 Mcps
Sampling rate	1 sample/chip
Power control	ON, both inner and outer loops
BLER target	10%
Rake finger allocation	Known delays
Maximum number of allocated Rake fingers	2 Rx ant.: 4·2·(# of Tx ant.) 4 Rx ant.: 2·4·(# of Tx ant.)
Channel estimation	Estimated from DPCH
Signal-to-interference est.	Estimated from DPCH

TABLE III.
SERVICE RELATED PARAMETERS

Data rate DTCH/DCCH (Dedicated Traffic Channel / Dedicated Control channel)	64 / 2.4 kb/s (circuit switched data)
Rate matching (DTCH)	16% repetition
Channel coding	Turbo 1/3
Interleaving	10 ms
Number of DPCHs	1
Spreading factor	256 / 16
Number of dedicated pilot / power control / transport format combination indicator (bits / slot)	6 / 2 / 2
Power ratio of DPCH/DPDCH	-5.46 dB

III. LINK PERFORMANCE

This section presents the link performance of different antenna configurations in a single user scenario in which the intra-cell interference is modeled as AWGN (Additive White Gaussian Noise). The measure for comparison is the average receive/transmit (Rx/Tx) E_b/N_0 (bit energy / total noise power density) that is required to obtain the desired Quality of Service (QoS) (10% block error rate). Average Rx E_b/N_0 per receiving antenna reflects to the amount of intra-cell interference that the other users experience from the given UE. Thus it is a measure of uplink capacity, i.e., the maximum number of users that can be supported by a single cell. Required Tx E_b/N_0 illustrates the average transmit power of the given UE and it corresponds to the amount of interference that the given UE generates to the other cells. Tx E_b/N_0 can be used as a performance measure also when comparing the UL coverage of different antenna configurations.

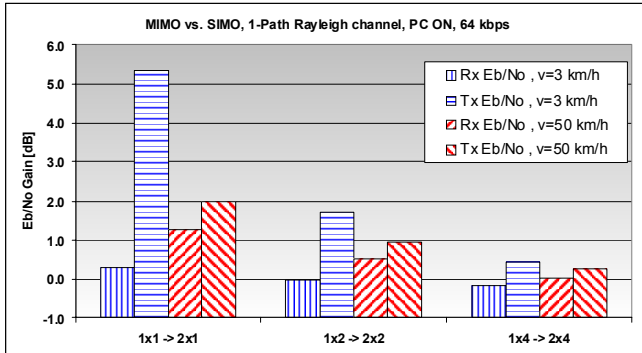


Figure 2. Transmit diversity gain with different number of receive antennas, flat fading channel. ($v=3, 50$ km/h).

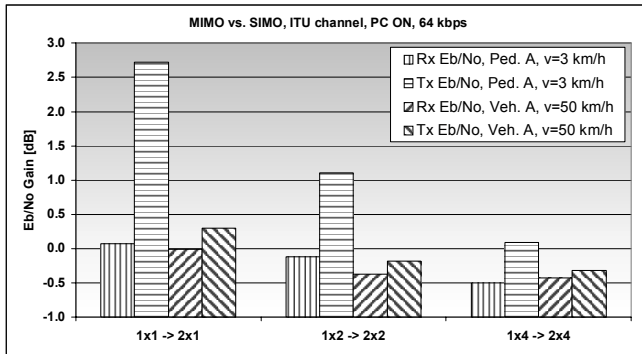


Figure 3. Transmit diversity gain with different number of receive antennas, mod. ITU Ped. A ($v=3$ km/h) and mod. ITU Veh A ($v=50$ km/h) channels.

Fig. 2 shows the uplink performance gain of the chosen MIMO scheme in a flat fading channel with 1, 2 and 4 receive antennas. In power controlled UL the gain is seen as reduced average transmit power of the mobile terminal rather than in the received Rx E_b/N_0 level. Results show that in all cases the gain of the 2nd transmit antenna decreases quite rapidly as the number of receiving antennas increases. Increasing the speed of mobile from 3 km/h to 50 km/h has the same effect: high degree of diversity is already available in the system (through the channel coding and interleaving). Thus, the additional antenna diversity gain is smaller. Fig. 3 illustrates the same results in modified ITU Pedestrian A and Vehicular A radio

channels. It can be seen from Fig. 3 that the additional antenna diversity gain is getting smaller when the amount of multipath diversity is increased. In a typical uplink case with Rx diversity the gain at the transmitter side is about 1 dB (mod ITU Ped. A channel, $v=3$ km/h). Introducing MIMO in multipath channels leads to some loss in Rx E_b/N_0 performance ($1x2 \Rightarrow 2x2$). $2x4$ MIMO performs about 0.5 dB worse than receive diversity with 4 antennas ($1x4$ case). Main reason for the above results is that errors in channel estimation increase with increased number of diversity branches. Since the uplink capacity is directly proportional to the required Rx E_b/N_0 per antenna, it is evident that receive diversity with 4 antennas is a good choice for optimizing user data throughput in uplink.

IV. ISOLATED CELL CAPACITY

Isolated cell capacity was studied by means of a multi-user simulation in which the intra-cell interference is being modeled as actual multiple-access interference. In the simulations each user meets the defined QoS target (10% BLER) by having both inner and outer loop power control loops running. It should be noted that capacity evaluation is valid only in the isolated cell environment since the system level issues such as soft handover and inter-cell interference are not taken into account.

Fig. 4 presents the simulated noise rise curves for the considered SIMO and MIMO transmission schemes. The presented noise rise is given as a function of number of 64 kb/s users in an isolated cell. The noise rise of 3.0 dB corresponds to a 50% load factor, and the noise rise of 6.0 dB to a 75% load factor. We can notice from Fig. 4a) that the usage of the 2nd transmit antenna does not increase the isolated cell capacity in a one path Rayleigh channel ($v=3$ km/h). UL power control (PC) translates into an additional diversity gain provided by the MIMO transmission. However, this gain is visible only at the transmit (UE) side. For that reason the Rx E_b/N_0 requirement is not improved using MIMO and thereby the isolated cell capacity is approximately the same as in the SIMO case. Doubling the number of receiving antennas will double the isolated cell capacity in both SIMO and MIMO cases.

Fig. 4b) illustrates again one-path Rayleigh channel, but now the speed of mobile is 50 km/h. This is the only simulated case in which the usage of the 2nd transmit antenna provides some capacity gain in an isolated cell (about 10% with 2 receive antennas). The explanation for that is the non-ideal behavior of the power control. UE speed of 50 km/h is relatively high velocity from the PC point of view and it cannot fully compensate the fast fading. Since the MIMO has smaller variance in the received power than SIMO (due to the higher number of diversity branches) the "slowness" of the PC loop does not impact as much in the MIMO case.

Fig. 4c) represents the simulated noise rise in the modified ITU Pedestrian A channels ($v=3$ km/h). We can notice that the usage of the 2nd transmit antenna will actually degrade the isolated cell capacity especially in case of 4 receiving antennas. The reason for that is that the estimation errors (combining losses) are larger in the MIMO case since the DPCH must be transmitted via both antennas (pilot power per antenna is 3 dB smaller than in SIMO case). Performance degradation caused by the 2nd transmit antenna increases as, number of multipath

components in the radio channel increases. The degradation is due to the fact that the SNR (signal-to-noise ratio) per Rake finger decreases as the number of multi-path components increases. This is seen in Fig. 4d) which represents noise rise curves in the modified Vehicular A channel ($v=50$ km/h). We can notice that the isolated cell capacity of the SISO case is almost 10% better than that of the MIMO case. Estimation errors increase also as the number of Tx/Rx antennas and/or the speed of mobile increase. In addition, the considered service has also an impact on the estimation errors: for the high data rate the power of the dedicated pilot channel can be increased without significant increase of the pilot overhead and the estimation errors will be accordingly smaller.

As a summary, it can be said that the doubling of the number of receiving antennas will almost double the isolated cell capacity, whereas the increase of the number of transmit antennas will mostly have negative impact on the isolated cell capacity. Moreover, the capacity gain from doubling the Rx antenna count from two to four may be even slightly pessimistic in the Vehicular A channel (Fig. 4d) since eight Rake fingers per transmit antenna were allocated in all cases. Accordingly, the 4-antenna receivers exploit only a half of the multipath components employed by the 2-antenna receivers (see Table II). Thus the 4-antenna receiver does not exploit all the signal energy that could be used.

V. UPLINK COVERAGE

Coverage calculation is based on the average transmitted E_b/N_0 (dB) values. Fig. 5a) presents the relative coverage gain of three different antenna configurations (2x2, 1x4 and 2x4) in a single user scenario. The reference case is the 1x2 case. We can notice in Fig. 5a) that the 2x2 MIMO case can support up to 2 dB higher path loss than the SISO scheme with 2 receiving antennas. The coverage gain of MIMO decreases as the degree of time, multipath and/or Rx diversity in the radio channel increases. The gain can be also negative (mod. ITU Vehicular A channel, $v=50$ km/h) when the additional estimation errors are larger than the additional diversity gain provided by the 2nd transmit antenna. One finding is also the fact that for the same degree of diversity the 1x4 antenna constellation is 3 dB better than the 2x2 constellation.

Figs. 5b) and 5c) illustrate examples of coverage evaluation in the modified Pedestrian A channel with the UE speed of 3 km/h. Fig. 5b) compares the 2x2 configuration to the 1x2 case and Fig. 5c) the 1x4 constellation to the 1x2 case. We can notice that in the 1x4 case the coverage gain increases with the increasing UL load. The total coverage gain is obtained by taking both reduced Tx E_b/N_0 and reduced noise rise into account. This means that e.g., having 30 simultaneous 64 kb/s users in the sector of interest the 2x2 case provides 1.3 dB better coverage compared to 1x2 case whereas the 1x4 case can provide as large gain as 6.6 dB (2.3 dB + 4.3 dB). Table IV gives the total coverage gain figures for the considered SISO and MIMO approaches in the mod. ITU Pedestrian A channel ($v=3$ km/h).

As a conclusion from the uplink coverage studies it can be stated that increasing the number of receive antennas is more

feasible than employing the MIMO concept. The coverage improvement of 4 Rx antennas over 2 Rx antennas is significant. Those operators, which do not have existing base station infrastructure can exploit this gain effectively since fewer base station sites are needed for the initial network roll-out. Moreover, as the data services get more popular this gain can be also realized as improved coverage for medium or high data rates.

TABLE IV.
COVERAGE GAIN IN MOD ITU PEDESTRIAN A CHANNEL ($v=3$ km/h)

	Relative coverage gain vs. 1x2 configuration (dB)			
	1 user	10 users	20 users	30 users
2x2	1.3	1.3	1.3	1.3
1x4	4.3	4.8	5.5	6.6

VI. CONCLUSIONS

The results showed that transmit diversity in UL does not increase the isolated cell capacity. The reason for that is that additional diversity does not decrease the average received E_b/N_0 value in the power controlled system. MIMO transmission will provide some gain (up to 2 dB) for UL coverage and UE battery consumption. Doubling the number of receiving antennas is much more effective than introducing the 2nd transmit antenna. The 4-antenna receive diversity approach can almost double the isolated cell capacity compared to the 2-antenna receiver and provide 3-6 dB better coverage than the 2x2 MIMO case. The coverage gain depends strongly on the UL load.

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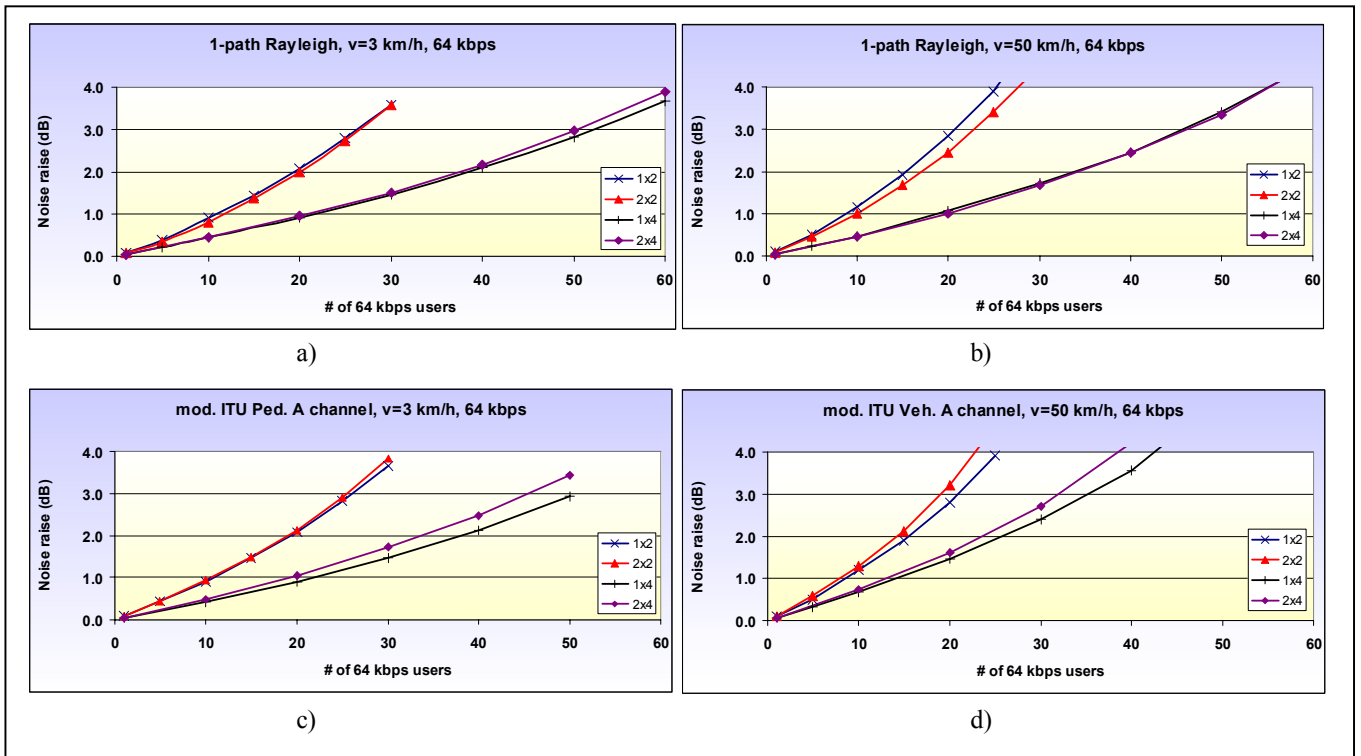


Figure 4. UL noise rise, SIMO and MIMO, 64 kb/s.

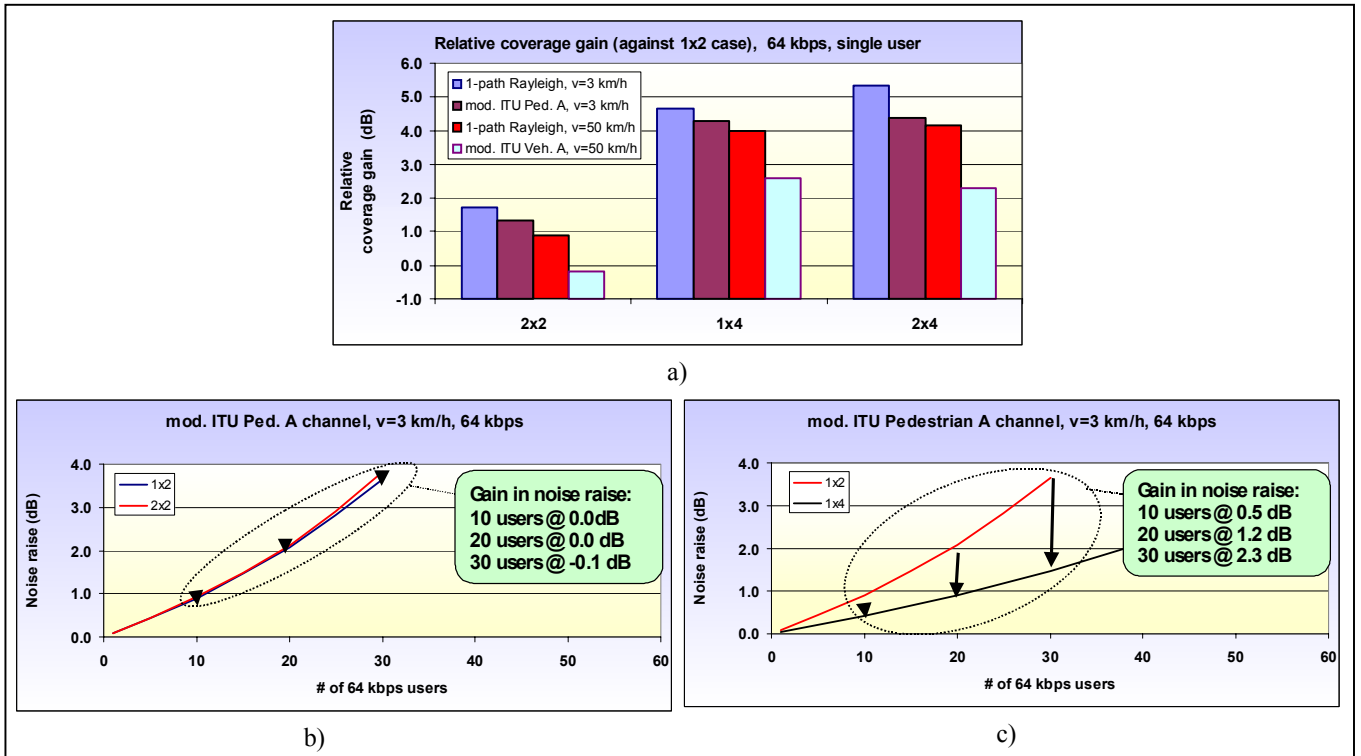


Figure 5. UL coverage, SIMO and MIMO, 64 kb/s.

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- III CONVENTIONAL CHANNEL ESTIMATION
- IV PROPOSED PA-DDCE SCHEME
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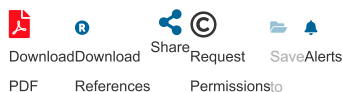
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By using the multiple antennas at both the transmitter and the receiver, the orthogonal frequency division multiplexing (OFDM) system can significantly increase the channel capacity, which makes it to be an attractive candidate for the high-data-rate wireless applications. Such systems are called the multiple-input multiple-output (MIMO) OFDM systems. To correctly demodulate the received data symbols, the channel estimation plays an important role to the system performance greatly. In practice the channel state information (CSI) is estimated from known pilot symbols and subsequently used for the data detection. Therefore the system performance depends on the quality of channel estimation and hence the number of pilot symbols. Estimation of the MIMO fading channel is a major challenge for the multiple-antenna wireless communication systems. When the number of antennas increases, the accurate channel estimation becomes more difficult due to the increase parameters to be estimated. In the conventional channel estimation, the pilots should be orthogonal to avoid the inter-antenna interference.

For the MIMO-OFDM systems with Frequency Domain Multiplexed (FDMed) pilots, the training pilots transmitted from different antennas occupy different sub-carriers exclusively within the same OFDM symbol duration. However, this method radically increases the pilot overhead when the number of the transmit antennas increase, which diminishes the system spectrum efficiency. The semi-blind and blind channel estimators [1] [2] have been proposed to reduce the pilot overhead; however, these methods are not practical due to their poor performance and high complexity. Motivated by this fact, a joint pilot-aided and blind decision-directed channel estimation (PA-DDCE) scheme is proposed in this paper to achieve a good tradeoff between the pilots overhead and the system performance. In the previous literatures [3] [4], the decision-directed channel estimation is proposed to support the high terminal velocities. The contribution of our proposed scheme is to design a pilot structure for the MIMO-OFDM system which has a relative small pilot overhead but still guarantees the accuracy of the channel estimation. At first the data symbols with the orthogonal format are transmitted in the middle of any two adjacent pilot sub-carriers. After that with the knowledge of the modulation alphabet of the data symbols, the blind hard decision on the symbols can be made and then the CSI estimate performance can be improved with slight increase of the pilot overhead versus the conventional orthogonal pilot structure. The pilot structure of the PA-DDCE scheme with two alternative orthogonal data formats have also been proposed in this paper, they are named as null-data format and space frequency block code (SFBC)-data [5] format.

The rest of the paper is organized as follows. Section II introduces the MIMO-OFDM system model and presents the conventional pilots-aided channel estimation in the MIMO-OFDM system. Then section III proposes our PA-DDCE scheme with two orthogonal data formats. The simulation results are given in section IV and finally we conclude our paper in section V.

SECTION II**THE MIMO-OFDM SYSTEM**

JOINT PILOT-AIDED AND BLIND DECISION-DIRECTED CHANNEL ESTIMATION FOR MIMO-OFDM SYSTEM

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ABSTRACT

In the conventional pilot-symbol-aided channel estimation for the MIMO-OFDM system, the pilots should be orthogonal in frequency, time, code, or spatial domain to mitigate the inter-antenna interference. In the frequency domain multiplexed (FDMed) pattern, the training pilots transmitted from different antennas conventionally occupy different sub-carriers exclusively within the same OFDM symbol duration. However, this method dramatically increases the pilot overhead as the number of the transmit antennas increases. In this paper a joint pilot-aided and blind decision-directed channel estimation (PA-DDCE) scheme is proposed to achieve a good tradeoff between the pilots overhead and the system performance. The proposed scheme utilizes the initial estimated channel obtained by the conventional channel estimation to detect the transmitted data symbol in the middle sub-carrier between any two adjacent pilot sub-carriers. Then these symbols are regarded as the "extra pilot symbols" and are applied to estimate the channel on the whole frequency band. From our simulation results the proposed PA-DDCE scheme can outperform the conventional channel estimation significantly in both the BER and the throughput performances.

I INTRODUCTION

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The rest of the paper is organized as follows. Section II introduces the MIMO-OFDM system model and presents the conventional pilots-aided channel estimation in the MIMO-OFDM system. Then section III proposes our PA-DDCE scheme with two orthogonal data formats. The simulation results are given in section IV and finally we conclude our paper in section V.

II THE MIMO-OFDM SYSTEM

An OFDM system with two transmitters and two receivers is shown in Fig.1.

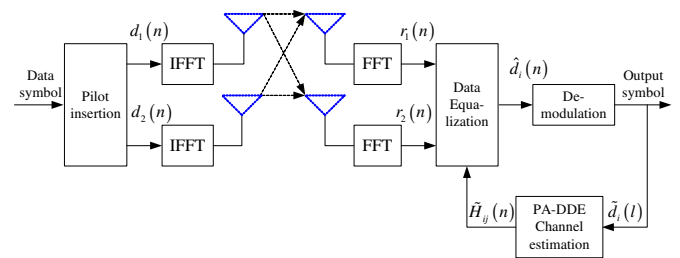


Figure 1: MIMO-OFDM system with PA-DDCE scheme

At one OFDM symbol duration, the transmitted symbols $\{d_i(n) : n = 0, 1, \dots, N-1\}$, for transmit antenna $i = 1, 2$, inserted with pilots on each antenna form an OFDM block, where N is the total number of OFDM sub-carriers. The two transmit antennas simultaneously transmit the OFDM signals modulated by inverse fast Fourier transform (IFFT). At the receiver side, the signal demodulated by fast Fourier transform (FFT) can be expressed as

$$r_j(n) = \sum_{i=1}^2 H_{ij}(n) d_i(n) + w_j(n), \quad (1)$$

where $H_{ij}(n)$ is the channel frequency response for the n th sub-carrier of the OFDM symbol, corresponding to the i th transmit antenna and the j th receive antenna. $w_j(n)$ denotes the complex additive white Gaussian noise (AWGN) with zeros mean and variance σ_n^2 , on the j th receive antenna.

III CONVENTIONAL CHANNEL ESTIMATION

In the conventional channel estimation with FDMed pilots of the MIMO-OFDM system, to keep the pilot orthogonality, a training pilot from one antenna is transmitted at a sub-carrier while the remaining transmit antennas are idle at the same sub-carrier, which is shown in Fig.2, where p_1 and p_2 are pilot symbols at the first and the second antenna respectively.

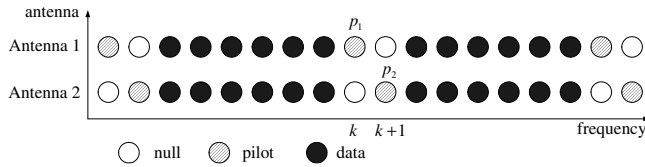


Figure 2: Pilot structure of the conventional channel estimation

For the MIMO-OFDM system with two transmit antennas and two receive antennas, by using Least Square (LS) method [6] the estimated channel frequency response on the pilot sub-carrier k and $k+1$ are given by

$$\begin{aligned} \hat{H}_{11}(k) &= r_1(k)/p_1(k) \\ \hat{H}_{12}(k) &= r_2(k)/p_1(k) \\ \hat{H}_{21}(k+1) &= r_1(k+1)/p_2(k+1) \\ \hat{H}_{22}(k+1) &= r_2(k+1)/p_2(k+1). \end{aligned} \quad (2)$$

The final estimated channel frequency response on the whole band sub-carriers is given by

$$\hat{H}_{ij}(n) = \text{interp}(\hat{H}_{ij}(k)), \quad (3)$$

where

- k pilot frequency subscript (sub-carrier indices)
- p_i transmitted pilot on i th transmitter antenna
- H estimated channel frequency response
- $\text{interp}()$ interpolation.

The performance of the conventional channel estimation for the MIMO-OFDM system greatly depends on the pilot number.

With more pilot symbols, the pilot spacing could be smaller thus implies less interpolation error. However, the system spectrum efficiency is diminished and the system throughput is decreased due to the larger pilot overhead. To overcome this problem, we propose a novel PA-DDCE scheme with two alternative pilot structures in the following section.

IV PROPOSED PA-DDCE SCHEME

In this section, the basic principle of the PA-DDCE scheme is introduced at first. Then we study the proposed scheme for two kinds of the orthogonal data symbol formats, namely null-data format and SFBC-data format.

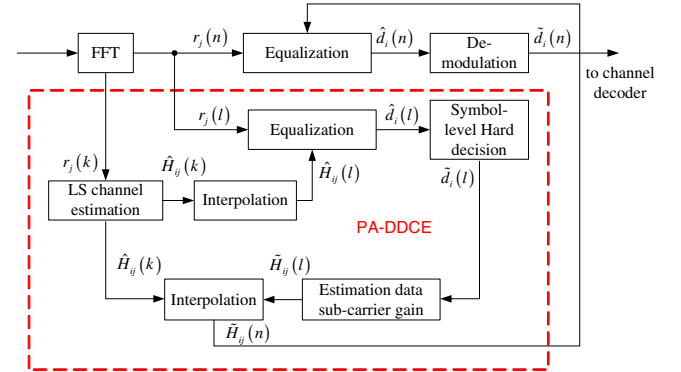


Figure 3: Principle of PA-DDCE scheme

The principle of the PA-DDCE scheme is illustrated in Fig.3, and the algorithm procedure is described as followed. Step 1: LS channel estimation to calculate the estimated channel frequency response $\hat{H}_{ij}(k)$ on the pilot sub-carriers. Step 2: Linear or other interpolation methods to obtain the estimated channel frequency response on the non-pilot sub-carriers $\hat{H}_{ij}(l)$, where l is defined as the middle sub-carriers between any two adjacent pilot sub-carriers. Step 3: Equalize the received data symbols $r_i(l)$ with the previous estimated channel frequency response $\hat{H}_{ij}(l)$ in step 2 and get the equalized data symbol $\hat{d}_i(l)$. Step 4: Hard decision of the equalized data symbol by demodulation and re-modulation, finally obtain the constellation points $\tilde{d}_i(l)$. Step 5: Calculate the new data sub-carrier channel frequency response $\tilde{H}_{ij}(l)$ by using the detected data symbols. Step 6: Linear or other interpolation with both $\hat{H}_{ij}(k)$ and new $\tilde{H}_{ij}(l)$ updated in step 5 to obtain the final estimated channel frequency response $\tilde{H}_{ij}(n)$.

In the followed paragraphs, the detailed descriptions of the step 3 and step 5 in the PA-DDCE scheme will be presented with regard to the null-data format and the SFBC-data format.

A PA-DDCE Scheme with Null-Data Format

The pilot structure of the PA-DDCE scheme with null-data format is shown in Fig.4. The difference to the conventional one is that in the middle of any two adjacent pilot symbols, a null-data format is inserted. When the data symbol d_1 on the sub-carrier l is transmitted at the first antenna, the second antenna is idle.

Vice versa, when the data symbol d_2 on the sub-carrier $l + 1$ is transmitted at the second antenna, the first antenna is idle.

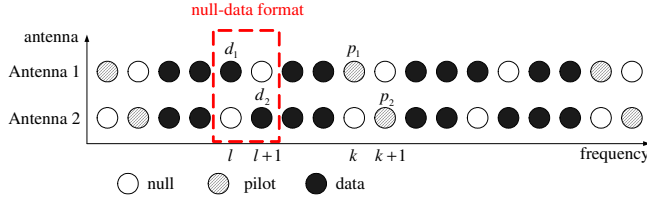


Figure 4: Pilot structure of PA-DDCE scheme with null-data format

Both the received data symbols on the sub-carrier l and $l + 1$ can be shown by

$$\mathbf{r} = \mathbf{H}\mathbf{d}_n + \mathbf{w}, \quad (4)$$

where we have

$$\mathbf{r} = \begin{bmatrix} r_1(l) & r_1(l+1) \\ r_2(l) & r_2(l+1) \end{bmatrix}, \quad (5)$$

$$\mathbf{H} = \begin{bmatrix} H_{11}(l) & H_{21}(l+1) \\ H_{12}(l) & H_{22}(l+1) \end{bmatrix}, \quad (6)$$

and

$$\mathbf{d}_n = \begin{bmatrix} d_1(l) & 0 \\ 0 & d_2(l+1) \end{bmatrix}. \quad (7)$$

The receiver adopts the two-branch maximal-ratio receiver combining (MRRC) and the transmitted data symbol $\hat{d}_1(l)$ and $\hat{d}_2(l+1)$ on the sub-carrier l and $l + 1$ are computed by zero forcing (ZF) equalizer, i.e.,

$$\hat{d}_1(l) = \frac{\hat{H}_{11}^*(l) r_1(l) + \hat{H}_{12}^*(l) r_2(l)}{|\hat{H}_{11}(l)|^2 + |\hat{H}_{12}(l)|^2}, \quad (8)$$

and

$$\hat{d}_2(l+1) = \frac{\hat{H}_{21}^*(l+1) r_1(l+1) + \hat{H}_{22}^*(l+1) r_2(l+1)}{|\hat{H}_{21}(l+1)|^2 + |\hat{H}_{22}(l+1)|^2}, \quad (9)$$

where $()^*$ denotes the complex conjugation and $|\cdot|$ is the absolute value. $\hat{H}_{ij}(l)$ and $\hat{H}_{ij}(l+1)$ are the estimated channel frequency response by the conventional channel estimation with the linear or other interpolation methods, which are obtained in step 2.

With the pre-known modulation alphabets, the symbol-level hard-decision is made on the transmitted data symbols $\hat{d}_1(l)$ and $\hat{d}_2(l+1)$ calculated in step 3. Assuming the error-free of the symbol hard decision, the re-modulated data symbol $\tilde{d}_1(l)$ and $\tilde{d}_1(l+1)$ on the constellation points can be regarded as extra “pilot symbols”, which means the pilot spacing is reduced to 1/2 of that in the conventional channel estimation. A new channel estimation on the l th and $l + 1$ th sub-carriers can thus be performed with these new “pilot symbols”, which leads to

$$\tilde{\mathbf{H}}_n = (\tilde{\mathbf{d}}_n^H \tilde{\mathbf{d}}_n)^{-1} \tilde{\mathbf{d}}_n^H \mathbf{r}, \quad (10)$$

for the new channel estimate, where we have

$$\tilde{\mathbf{H}}_n = \begin{bmatrix} \tilde{H}_{11}(l) & \tilde{H}_{21}(l+1) \\ \tilde{H}_{12}(l) & \tilde{H}_{22}(l+1) \end{bmatrix}, \quad (11)$$

and

$$\tilde{\mathbf{d}}_n = \begin{bmatrix} \tilde{d}_1(l) & 0 \\ 0 & \tilde{d}_2(l+1) \end{bmatrix}. \quad (12)$$

The update of the estimated channel frequency response is carried out by processing the previous estimated channel frequency response and the new ones through the interpolation. Thus the final enhanced estimated channel frequency response is given by

$$\hat{H}_{ij}(n) = \text{interp}(\hat{H}_{ij}(k), \hat{H}_{ij}(l)). \quad (13)$$

B PA-DDCE Scheme with SFBC-Data Format

The pilot structure of the PA-DDCE scheme with the SFBC-data format is shown in Fig.5.

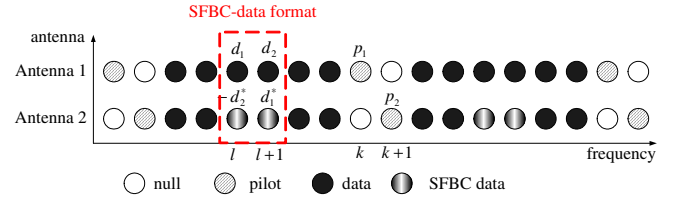


Figure 5: Pilot structure of PA-DDCE scheme with SFBC-data format

In the middle of any two adjacent pilot sub-carriers, the data symbol d_1 and d_2 on the sub-carriers l and $l + 1$ are encoded by the SFBC. Then the received signals on the sub-carrier l and $l + 1$ can be written as

$$\mathbf{r} = \mathbf{H}\mathbf{d}_s + \mathbf{w}, \quad (14)$$

where

$$\mathbf{d}_s = \begin{bmatrix} d_1(l) & d_2(l+1) \\ -d_2^*(l) & d_1^*(l+1) \end{bmatrix}. \quad (15)$$

Assuming the channel fluctuation on the adjacent sub-carriers is constant, i.e., $H_{ij}(l) = H_{ij}(l+1)$. The combined received signals are equivalent to that of the four-branch MRRC, thus the transmitted data symbols $\hat{d}_1(l)$ and $\hat{d}_2(l+1)$ on the sub-carrier l and $l + 1$ are computed by ZF equalizer, i.e.,

$$\begin{bmatrix} \hat{d}_1(l) \\ \hat{d}_2(l+1) \end{bmatrix} = \alpha \begin{bmatrix} \hat{H}_{11}^*(l) & \hat{H}_{21}(l) & \hat{H}_{12}^*(l) & \hat{H}_{22}(l) \\ \hat{H}_{21}^*(l) & -\hat{H}_{11}(l) & \hat{H}_{22}^*(l) & -\hat{H}_{12}(l) \end{bmatrix} \times \begin{bmatrix} r_1(l) \\ r_1^*(l+1) \\ r_2(l) \\ r_2^*(l+1) \end{bmatrix}, \quad (16)$$

where

$$\alpha = \frac{1}{\left| \hat{H}_{11}(l) \right|^2 + \left| \hat{H}_{21}(l) \right|^2 + \left| \hat{H}_{12}(l) \right|^2 + \left| \hat{H}_{22}(l) \right|^2}, \quad (17)$$

$\hat{H}_{ij}(l)$ are the estimated channel frequency response by the conventional channel estimation with the linear interpolation or other interpolation method, which are obtained in step 2.

The equalized symbol $\hat{d}_1(l)$ and $\hat{d}_2(l+1)$ are then demodulated and re-modulated to $\tilde{d}_1(l)$ and $\tilde{d}_1(l+1)$ respectively by the symbol-level hard-decision which are regarded as extra “pilot symbols” and can be used to compute the channel frequency response at the sub-carriers l and $l+1$:

$$\tilde{\mathbf{H}}_s = \left(\tilde{\mathbf{d}}_s^H \tilde{\mathbf{d}}_s \right)^{-1} \tilde{\mathbf{d}}_s^H \mathbf{r}, \quad (18)$$

where

$$\tilde{\mathbf{H}}_s = \begin{bmatrix} \tilde{H}_{11}(l) & \tilde{H}_{21}(l) \\ \tilde{H}_{12}(l) & \tilde{H}_{22}(l) \end{bmatrix}, \quad (19)$$

and

$$\tilde{\mathbf{d}}_s = \begin{bmatrix} \tilde{d}_1(l) & \tilde{d}_2(l+1) \\ -\tilde{d}_2^*(l) & \tilde{d}_1^*(l+1) \end{bmatrix}. \quad (20)$$

The final enhanced estimated channel frequency response can be written as

$$\hat{H}_{ij}(n) = \text{interp} \left(\hat{H}_{ij}(k), \tilde{H}_{ij}(l) \right). \quad (21)$$

V NUMERICAL RESULTS

Table 1: simulation parameters

Bandwidth	5 MHz
Number of Sub-carrier	1024
Channel Model	ITU-Pedestrian B
Channel Coding	Turbo Coding, Rate 1/2
Data Modulation	QPSK
Antenna Configuration	2Tx, 2Rx
Pilot Modulation	QPSK
Data Detection	MMSE
Repeat Factor (RPF)	24, 36
Channel Estimation	Ideal, Conventional (conv), PA-DDCE with null data (dd-null), PA-DDCE with SFBC data (dd-SFBC).
Pilot Overhead	8.4 %
Interpolation	Linear

In this section, the performance of the proposed PA-DDCE scheme is measured by the bit error rate (BER) and the system throughput, which are compared with those of the conventional channel estimation. The simulation specifications are presented in Table I. The repeat factor (RPF) means the sub-carrier spacing between the adjacent pilots. For fair comparison, the pilot overhead is assumed to be the same, thus for conventional channel estimation the RPF is 24 while for PA-DDCE

scheme it is 36. In other words, we constrain the useful data sub-carriers the same. The channels are quasi-static Rayleigh faded over a coding frame and independent among different frames. We also assume the channels between the transmit antennas and the receive antennas are independent with each other. Here the ITU-PB channel [7] refers to a 6 taps Rayleigh fading channel. To detect the data symbols the well-known linear minimum mean square error (MMSE) [8] solution is applied for the MIMO-OFDM system in our simulation.

A Estimated channel accuracy comparison

Fig.6 plots the estimated channel gain $|\hat{H}|^2$ obtained from both the conventional estimation and the proposed PA-DDCE scheme with SFBC-data format in comparison with the ideal channel gain $|H|^2$. As it is shown when SNR is 20dB the estimated channel gain in the middle of any two adjacent pilot sub-carriers computed from the extra “pilot symbols” by PA-DDCE scheme can improve the total estimated channel quality dramatically.

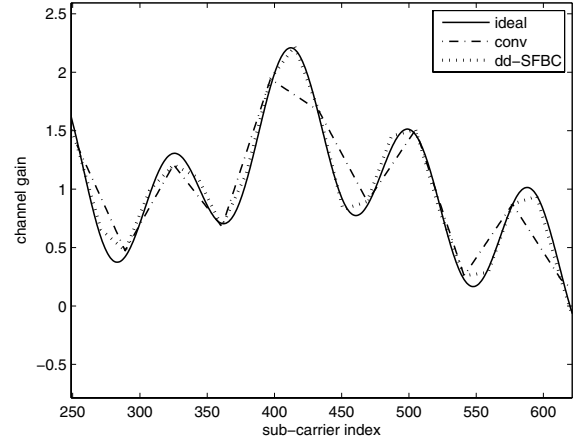


Figure 6: Estimated channel comparison under SNR = 20dB

The Mean Square Error (MSE) of the estimated channel by using different channel estimations is depicted in Fig.7. Applying the PA-DDCE scheme, it can be observed that the MSE performance is significantly improved compared with the conventional channel estimation. Under high SNR, the improvements become larger due to the higher reliability of the symbol-level hard decision. Further more, the PA-DDCE scheme with the SFBC-data format works even better than that with the null-data format. The reason is that for the SFBC-data format, the received extra “pilot symbols” are combined as the four-branch MRRC while for the null-data format they are combined as the two-branch MRRC.

B BER and throughput performance

The channel estimation performance is mainly dominated by two factors: the estimation error on the pilot sub-carriers and the interpolation error on the none-pilot sub-carriers. When the

pilots spacing is larger than the channel coherent bandwidth, the interpolation error becomes the major problem affecting the channel estimation. Thus with fixed small pilot overhead, the conventional channel estimation can not achieve the required system throughput performance. When the channel is ideal, to guarantee the same system spectrum, we assume the pilot overhead is still 8.4%, i.e. RPF of 24. The results also show the proposed PA-DDCE scheme outperforms the conventional channel estimation significantly. For example, when the SNR is 12.5dB, the system throughput gains are 0.5 Mbps and 1 Mbps for the null-data format and the SFBC-data format respectively.

VI CONCLUSIONS

For the high-data-rate wireless applications especially utilizing the MIMO-OFDM technology as the air interface, the number of the channel parameters used for the data detection increased dramatically. To achieve the channel estimation quality while still maintaining the high spectrum efficiency, a novel PA-DDCE scheme for the MIMO-OFDM system is presented

in this paper. We also proposed two alternative pilot structures with the null-data format and the SFBC-data format respectively to mitigate the inter-antenna interference for the PA-DDCE scheme. From our simulation results the PA-DDCE scheme can outperform the conventional channel estimation significantly respect to both the BER and the throughput performances. The only disadvantage is a slight but acceptable complexity on the receiver side.

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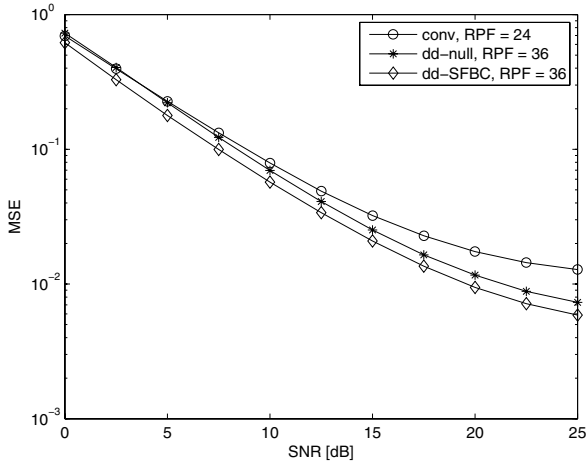


Figure 7: MSE performance of QPSK 1/2 for ITU-PB channel

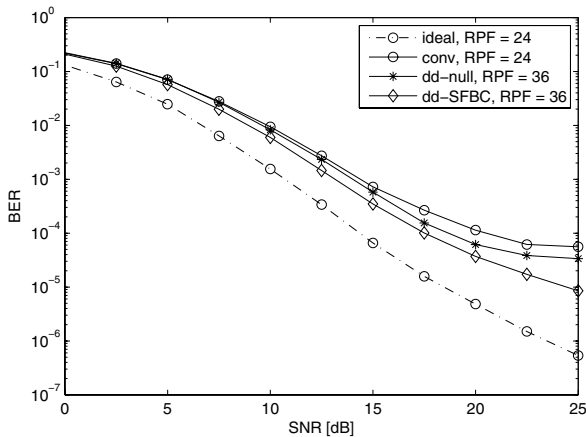


Figure 8: BER performance of QPSK 1/2 for ITU-PB channel

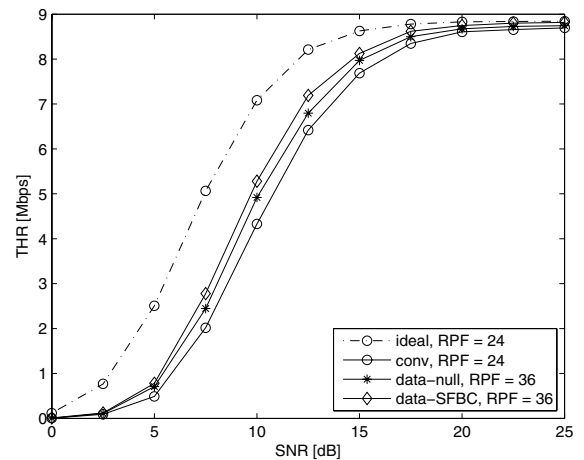


Figure 9: Throughput performance of QPSK 1/2 for ITU-PB channel